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Widebanding Techniques for VHF Antennas - I

by

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A straight wire antenna of fixed length is narrow band due to the rapid variations in its input impedance, and to a lesser degree, in its radiation pattern. A wire antenna is, however, lightweight, easy to fabricate and simple to mount on any platform. To make a wire antenna broadband, we need resistive and/or reactive loading along its length. In this report, we consider monopole antennas loaded with (a) modified Wu-King resistivity profile, and (b) complex lumped loads. A 1-m monopole antenna is considered over 30-90 MHz. Input impedance and radiation efficiency data for the two cases are generated by employing the moment method. Tuning networks are designed using the real frequency method of Carlin to maximize the power transfer from a 50 ohm source to the antenna. Overall performance of the antenna by taking into account the mismatch loss due to the tuning network and resistive loss due to the loading is studied.

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ABSTRACT

A straight wire antenna of fixed length is narrow band due to the rapid variations in its input impedance, and to a lesser degree, in its radiation pattern. A wire antenna is, however, lightweight, easy to fabricate and simple to mount on any platform. To make a wire antenna broadband, we need resistive and/or reactive loading along its length. In this report, we consider monopole antennas loaded with (a) modified Wu-King resistivity profile, and (b) complex lumped loads. A 1-m monopole antenna is considered over 30–90 MHz. Input impedance and radiation efficiency data for the two cases are generated by employing the moment method. Tuning networks are designed using the real frequency method of Carlin to maximize the power transfer from a 50 ohm source to the antenna. Overall performance of the antenna by taking into account the mismatch loss due to the tuning network and resistive loss due to the loading is studied.

I. INTRODUCTION

In the first phase of the project we have designed a broadband antenna to operate over the frequency range 30–90 MHz. A simple whip consisting of a 1m high resistively loaded monopole antenna was considered. A lossless tuning network was designed to render the input impedance as well as the radiation pattern of the antenna relatively insensitive to frequency. The antenna is cheap to fabricate, light-weight, and can to be easily mounted on moving vehicles or back-packs. FORTRAN programs were developed to facilitate computer aided design on a PC.

As is well known, a simple conducting wire antenna is extremely narrow band with respect to both its input impedance and radiation pattern. This is due primarily to a standing wave current that exists on the antenna. To make the wire antenna broadband, resistive coating can be employed so that current on the structure becomes predominantly traveling wave in nature. Wu and King [1] proposed a continuous resistive loading in the form of carbon coating to accomplish this. Compared to an unloaded antenna, a loaded antenna will be broadband, but will still exhibit excursions, albeit smaller, in its input impedance with respect to frequency. To reduce these impedance excursions even further, a tuning (or matching) network has to be designed. In this report, we present design procedure and performance results for a resitively loaded monopole terminated in an appropriate tuning network.

II. RESISTIVELY LOADED MONOPOLE

For a center fed dipole antenna of length $2h$ and radius a with its axis along the z axis, Wu and King [1] have suggested a continuously varying resistive (actually impedance) loading of the form

$$Z_i(z) = \frac{60\Psi}{(h-|z|)} \quad \text{ohms/m} \tag{1}$$

where

$$\Psi = 2\left[\sinh^{-1}\frac{h}{a} - \int_0^{2kh} \frac{1-\cos\sqrt{u^2+4k^2a^2}}{\sqrt{u^2+4k^2a^2}}\,du - j\int_0^{2kh}\frac{\sin\sqrt{u^2+4k^2a^2}}{\sqrt{u^2+4k^2a^2}}\,du\right] + \frac{j}{kh}(1-e^{-jkh}) \tag{2}$$

and k is the wavenumber $2\pi/\lambda$.

For a monopole of length $h = 1$ m, $a = 0.5$ cm, we have calculated Ψ at the geometric mean frequency of 51.96 MHz in the desired band of 30–90 MHz. For the above parameters, we obtain

$$Z_i(z) \approx \frac{554.4 - j115.2}{(h-z)} \quad \text{ohms/m}, \tag{3}$$

where the integrals in (2) have been approximated as the cosine and sine integrals, Cin (x) and Si (x) respectively.

Since a fixed reactance over the entire frequency band is difficult to realize, we shall ignore the reactive part of (3) and assume that the antenna is only resistively loaded. It has been demonstrated in [8] that Wu-King resistivity profile as given by the first part of (3) results in poor efficiencies, particularly at lower frequencies where the antenna is electrically short. A resistivity profile which is chosen to be 30 – 40% of the Wu-King profile was shown to result in higher efficiencies compared to the Wu-King profile, while at the same time producing a slowly varing input impedance. We have used a 31% Wu-King profile, i.e., one having a resitivity

$$R_i(z) = \frac{171.6}{(h-z)} = \frac{R_i(0)}{(1 - z/h)} \quad \text{ohms/m} \tag{4}$$

where $R_i(0) = 171.6/h = 0.31 \times 554.4/h$. Fig. 1 shows the variation of R_i as a function of distance from the feed point. The resistivity ranges from a small value at the feed point to large values as $z \longrightarrow h$.

The above resistivity could be realized in two ways. If a conducting cylinder of radius a is coated with a lossy film of thickness d and conductivity σ, it provides a DC resistance per unit length $R_i = \frac{1}{2\pi a \sigma d}$ for $d \ll a$. If carbon (industrial graphite) is used as the lossy material with $\sigma \approx 7 \times 10^4$ S/m [2] (skin depth = 0.347 mm at 30 MHz, 0.2 mm at 90 MHz), its thickness $d(z)$ to realize the resistivity profile in (4) can be obtained as

$$d(z) = 2.65(1 - z/h)\,\mu\text{m} \tag{5}$$

Alternately, one may use a discrete approximation to the continuous profile (4). Assuming that the antenna height h is subdivided into N equal segments of size Δz, a fixed resistance of value

$$R_n = \int_{(n-1)\Delta z}^{n\Delta z} R_i(z)\,dz = hR_i(0)\ln\left|\frac{N+1-n}{N-n}\right| \tag{6}$$

may be connected at the junction of n^{th} and $(n+1)^{\text{th}}$ segments, where segment numbering has been assumed to start at the base of the monopole towards its end. Table 1. shows the values of R_n for the 1 m high monopole using $N = 9$ segments.

We have used the latter approach in the numerical medeling of the loaded monopole antenna. With the monopole loaded as shown in Table 1., we have calculated its input impedance on an infinite perfectly conducting ground plane. The loaded monopole was modeled using the method of moments with triangular basis functions and point matching. Table 2. shows the input impedance of the monopole as a function of frequency over the desired band. For the sake of comparison, the input impedance of the unloaded monopole is also shown. Whereas the unloaded monopole has a widely varying input resistance and a reactance that changes from capacitive to inductive, the loaded monopole has a slowly varying resistance and a reactance that remains capacitive throughout the frequency band. This will result in a simpler tuning network for the loaded monopole. The loading of monopoles will however result in reduced radiation efficiency of the antenna.

Assuming that the base metal used in the unloaded monopole is copper, its total loss resistance R_{cu} is given by [3]

$$R_{\text{cu}} = \frac{h}{2\pi a}\sqrt{\frac{\omega\mu_o}{2\sigma}} = \frac{h}{a}\sqrt{\frac{f}{2\sigma}} \quad (f \text{ in MHz}, \sigma \text{ in } S/\text{m}).$$

At a frequency of 30 MHz, the 1m high, 1cm diameter monopole will have a loss resistance R_{cu} of 4.59×10^{-2} ohms. The radiation resistance of the same monopole at 30 MHz is 3.87 ohms from Table 2. (also $\approx 40\pi^2(h/\lambda)^2$ [3]). The radiation efficiency of the unloaded monopole is then 98.8% at 30 MHz. At higher frequencies it will be even better. We shall therefore regard the unloaded monopole as being lossless and assume that the losses on the loaded monopole are comprised entirely of the load resistors. The radiation efficiency η_{rad} of the loaded antenna is computed using

$$\eta_{\text{rad}} = 1 - \frac{P_{\text{loss}}}{P_{\text{in}}} = 1 - \frac{\sum_{n=1}^{N-1} |I_n|^2 R_n}{|I_o|^2 R_{\text{in}}} \tag{7}$$

where P_{loss} is the power loss due to resistive loading, P_{in} is the input power supplied to the antenna, I_o is the base current, and I_n is the current on the monopole flowing at the location of R_n. Table 3. shows the calculated input power, the loss power and the radiation efficiency of the monopole antenna. Compared to the unloaded case, it is seen that the efficiency at 30 MHz is reduced from 98.8% to about 6% in the loaded antenna. Clearly broadbanding is at the expense of lower radiation efficiency. Fig. 2 shows the efficiency of the antenna plotted as function of frequency.

Fig. 3 shows that relative power patterns of the loaded monopoles at 30, 60, and 90 MHz. The axis of the monopole corresponds to 0^o on the plot. The pattern is seen to be relatively insensitive to frequency. It is clear from Table 2. that the antenna cannot be ragarded yet as a broadband antenna as its input impedance vaires with frequency. For example, the Transducer Power Gain (TPG) defined as the ratio of the power tranferred to the antenna to the power available from a 50 ohm source shown in Fig. 3 is seen to vary significantly over the frequency band. To make the antenna broadband with respect to the source, an equalizing network inserted between the source and the antenna is to be designed. The goal of the equalizer, or the tuning network, is to provide a relatively flat TPG efficiency over the desired band. This is accomplished in the next section. It is to be borne in mind that an arbitrary TPG cannot be obtained without restricting the bandwidth and the TPG as exemplified by Fano [4].

III. DESIGN OF TUNING NETWORK

We shall design a tuning network for the antenna using the real frequency method of Carlin [5]. The method is a numerical one, and only uses real frequency load impedance data $Z_l(j\omega) = R_l(\omega) + jX_l(\omega)$. No model or analytical impedance function for the load nor an equalizer topology is necessary. Instead the Thevenin impedance of the equalizer as seen from the load (the antenna in our case) $Z_q(j\omega) = R_q(\omega) + jX_q(\omega)$ is described in a simple manner in terms of a number of real parameters, which, once chosen to optimize gain-bandwidth define the equalizer design. More details can be found in [5]. Fig. 4 shows the various impedances as described above. The key steps involved in the design of a *band-pass* equalizer are given below. The starting in the method is an expression for the Transducer Power Gain (TPG) expressed as

$$TPG = \frac{\text{Power to load}}{\text{Power available from generator}} = 1 - |\rho|^2 = \frac{4R_l(\omega)R_q(\omega)}{|Z_l(j\omega) + Z_q(j\omega)|^2}, \tag{8}$$

where ρ is the reflection coefficient defined as $\rho = (Z_q - Z_l^*)/(Z_q + Z_l^*)$. The real frequency load data $R_l(\omega), X_l(\omega)$ are given, and the unknown equalizer reactance $X_q(\omega)$ is obtainable through $R_q(\omega)$ through a Hilbert transform [5]. In the numerical procedure, the TPG is assumed to be approximately flat over the prescribed passband, and we seek to maximize the minimum passband gain T_o with respect to the unknowns in Z_q.

The real part $R_q(\omega)$ is represented as a number of straight line segments with frequency break points $\omega_o = 0 < \omega_1 < \omega_2 \ldots < \omega_n$, where the resistance is assumed to go to zero beyond the last frequency ω_n. The choice of ω_n beyond which $R_q(\omega) = 0$ depends on the roll-off desired and the other break points ω_k are assumed to be known. The equalizer resistance is then

$$R_q(\omega) = \sum_{k=1}^{n} a_k(\omega) r_k, \tag{9}$$

where r_k is the unknown resistance excursion of the k^{th} straight line segment between the break points ω_{k-1} and ω_k and

$$a_k(\omega) = \begin{cases} 0, & \omega \le \omega_{k-1}; \\ \dfrac{\omega - \omega_k}{\omega - \omega_{k-1}}, & \omega_{k-1} \le \omega \le \omega_k; \\ 1, & \omega \ge \omega_k. \end{cases} \tag{10}$$

Note from (10) that we have chosen the DC resistance $R_q(0) = 0$. This is valid since we are dealing with a bandpass equalizer. Since $R_q(\omega) = 0$, $\omega \ge \omega_n$, we have

$$\sum_{k=1}^{n} r_k = 0, \tag{11}$$

so that there are $(n-1)$ unknowns, i.e., $r_k; k = 1, 2 \ldots, n$. As a result of this representation for $R_q(\omega)$, the reactance function $X_q(\omega)$ is also representable as a linear combination of the same unknown resistive excursions r_k. Thus the exact reactance corresponding to $R_q(\omega)$ is given by

$$X_q(\omega) = \sum_{k=1}^{n} b_k(\omega) r_k. \tag{13}$$

For any frequency ω, the coefficients $b_k(\omega)$ can be expressesd in terms of the break point frequencies by the Hilbert tranform property as

$$b_k(\omega) = \frac{1}{(\omega_k - \omega_{k-1})\pi} \int_{\omega_{k-1}}^{\omega_k} \ln \left| \frac{y+\omega}{y-\omega} \right| dy$$

$$= \frac{1}{(\omega_k - \omega_{k-1})\pi} \Big\{ (\omega + \omega_k) \ln(\omega + \omega_k) - (\omega + \omega_{k-1}) \ln(\omega + \omega_{k-1})$$

$$+ (\omega - \omega_k) \ln |\omega - \omega_k| - (\omega - \omega_{k-1}) \ln |\omega - \omega_{k-1}| \Big\}. \qquad (13)$$

Starting from an initial guess (which could be arbitrary) for r_k and an initial value T_o for the TPG, optimization is performed on the variables r_k such that a flat TPG as determined by (8) is maintained over the desired frequency band. The procedure is repeated until the assumed gain T_o can no longer be increased (this will be indicated by the occurence of negative resistances for the equalizer network at some of the break points). Once the line segments describing $R_q(\omega)$ (and hence $Z_q(\omega)$) have been found, an appropriate rational approximation $\tilde{R}_q(\omega) \approx R_q(\omega)$ is needed for the purpose of realization. A suitable form for our purposes (i.e., with $R_q(0) = 0$) is

$$\tilde{R}_q(\omega) = \frac{A(\omega^2)}{B(\omega^2)} = \frac{\sum_{m=1}^{M} A_m \omega^{2m}}{\sum_{n=0}^{N} B_m \omega^{2n}}, \qquad (14)$$

where B_o is taken to be 1, and $N < M$ in order that the resistance eventually reaches zero for large frequencies. A second optimization is carried out to determine the unknowns A_m and B_n in (14) for a known $R_q(\omega)$. For physical realizability, the resistance as defined by (14) must be non-negative over all frequencies. Since the unknowns in an unconstrained optimization procedure can assume positive as well as negative values, non-negativeness is not guaranteed in (14). To achieve this we adopt the procedure of [6]. As an illustration, the denominator polynomial in (14) may be written in terms of a second polymonial $P_N(\omega)$

$$P_N(\omega) = 1 + x_1\omega + \ldots + x_N\omega^N$$

as

$$B(\omega^2) = 0.5[P_N^2(\omega) + P_N^2(-\omega)]$$

which will always be non-negative. The coefficients B_n may be easily related to the coefficients x_n as

$$B_1 = x_1 + 2x_2$$
$$\vdots$$
$$B_n = x_n^2 + 2\left(x_{2n} + \sum_{k=2}^{n} x_{k-1}x_{2n-k+1}\right)$$
$$B_N = x_N^2$$

Taking the antenna input impedance shown in Table 2 as the load, we have determined the resistive excursions to produce a TPG of 0.6 in the band 30 MHz $\leq f \leq$ 90 MHz. The values of the equalizer resistances using the straight line approximation of (9) are shown in Table 4 at the discrete frequencies ω_k. We have normalized all frequencies to 90 MHz so that $\omega = 1$ corresponds to 90 MHz. In the numerical procedure we have included four out-of-band break points at $\omega = 0, 1/9, 2/9$, and 1.5. The other break points correspond to 30, 40, 50, 60, 70, and 80 MHz. In the rational approximation, one term was used for the numerator polynomial and six for the denominator polynomial. The correponding resistance values using rational approximation are also shown in Table. 4. Fig. 5 shows the straight line and rational function resistances of the equalizer network.

Once the real part of a rational resistance function has been found, the imaginary part can be found using one of the algebraic methods [7]. We have used the Gewertz procedure for doing so. Given

$$\tilde{R}_q(\omega) = \frac{A(\omega^2)}{B(\omega^2)} = \left.\frac{m_1 m_2 - n_1 n_2}{m_2^2 - n_2^2}\right|_{s=j\omega},$$

we wish to find

$$Z_q(\omega) = \frac{P(s)}{Q(s)} = \frac{m_1 + n_1}{m_2 + n_2},$$

where m_1, m_2 are the even parts, respectively, of $P(s)$ and $Q(s)$, and n_1, n_2 the odd parts. First, $Q(s)$ is found by factoring $B(-s^2)$ and retaining the left half plane roots as per

$$Q(s)Q(-s)]_{s=j\omega} = m_2^2 - n_2^2]_{s=j\omega} = B(\omega^2)$$

Having found $Q(s)$ we need to find $P(s)$. We assume $P(s)$ to be a poynomial of the same degree as $Q(s)$ with undetermined coefficients. We then find $(m_1 m_2 - n_1 n_2)$ and equate, term by term, with the given $A(\omega^2)$ for $s^2 = -\omega^2$. This will yield the values for the undetermined coefficients in $P(s)$. For the resistance function shown in Table 4, we determine the equalizer impedance as

$$Z_q(s) = \frac{P(s)}{Q(s)} = \frac{400.64s + 705s^2 + 741.81s^3 + 283.61s^4}{0.447 + 1.62s + 3.36s^2 + 3.23s^3 + 2.62s^4 + s^5}. \tag{15}$$

The impedance in (15) can be realized as lossless tuning network with a resistive termination using the Darlington's procedure. Two different forms can be found depending on whether the expansion of $Z_q(s)$ is made about zero or infinity [7]. Fig. 6 shows the two lossless network realizations for the tuning network operating terminated in a 50 ohm source. With the tuning network connected to the antenna, the TPG of the system can be computed using (8) and is shown in Fig. 7. Also plotted is the TPG for the untuned case where the antenna is connected directly to the 50 ohm source. It is seen that the tuning network effectively compensates for the impedance variations of the dipole and provides a relatively flat gain over the desired band. It is also interesting to look at the input VSWR as seen from the 50 ohm generator. Fig. 8 shows the input VSWR as a function of frequency. It is seen that the VSWR is generally under 4.0 over most of the band resulting in power transmission of about 65% of power from the source. This is consistent with Fig. 7. It is to be borne in mind that even though the transducer power gain for the system is around 0.65, the actual power radiated by the antenna is less that 65% of the available power. This is because some power is lost in the loading resistors placed on the antenna. The actual power radiated by the antenna is proportional to $TPG \times \eta_{\mathrm{rad}}$ and ranges from 3.6% (0.65×5.54) of the available power at 30 MHz to about 18.4% (0.65×28.36) at 90 MHz.

We have attempted to design a matching network for the conducting monopole antenna without the resitive coating. This will be attractive owing to its potentially high radiation efficiency. It was found that a flat gain over the band could be maintained only at the expense of a relatively small TPG (of the order of 0.05), implyimg that the input VSWR with the matching network would be greater than 78. Although no power is dissipated by the antenna structure in this case, we see that most of the available power from the source is reflected back. Only about 5% of the available power is delivered to the antenna for radiation. At the lower frequencies, this is comparable to the previous case of resistively loaded antenna and does not present much advantage.

With the objective of increasing the antenna efficiency, while at the same time maintaining an appreciable TPG, we have tried other loadings on the antenna. The 1m monopole was loaded with two parallel RLC networks as shown in Fig. 9. In order for the load to actually affect the antenna impedance characteristics, the values of the elements and their locations must be chosen such that the overall impedance of the load becomes comparable to the input impedance of the antenna when excited from the load terminals. Figs. 10–14 show the variation of the input impedance of a conducting monopole of height 1m and radius 0.5cm as a function of the feed location. The point $z = 0$ corresponds to the base and the point $z = h$ to the end of the monopole. These figures should aid the decision on the placement of the loads and their values. To increase the radiation efficiency of the antenna at the lower frequencies, we need an inductive loads so that its electrical length is increased. However, an inductive load makes the input impedance of the monopole become more inductive for frequencies immediately beyond the first resonance (around 70 MHz for the 1m mmonopole). To reduce the impedance excursions, we need a capacitive load at higher frequencies. This will be furnished by a parallel RLC network if its resonance frequency is chosen to lie around the resonance frequency of the unloaded monopole. The resistance of the tank circuit is chosen to present a resonant resistance with the same order of magnitude as the real part of the input impedance of the monopole at the load terminals. Based on these arguments we have chosen the locations of the loads to be at 1/3m and 2/3m. The elements of the first tank circuit were chosen as $R_1 = 165$ ohms, $L_1 = 0.5\mu H$, $C_1 = 14.0pF$, and those of the second as $R_2 = 150$ ohms, $L_2 = 0.25\mu H$, and $C_2 = 20pF$. The first tank circuit has a resonance frequency at 60 MHz, whereas the second at 70 MHz. Table 5. shows the impedance of the loads, the input impedance of the antenna, and its radiation efficiency over 30–90 MHz. Compared to the multiple resistive loading case of Tables 2 and 3, it is seen that the doubly loaded monopole of Fig. 9 has much higher efficiency at the lower frequencies, while at the same time maintaining a slowly varying input impedance. We have designed a tuning network for this antenna based on the theory presented here. The equalizer impedance normalized to 50 ohms and

90 MHz is obtained as

$$Z_q(s) = \frac{P(s)}{Q(s)} = \frac{11.163s + 8.394s^2 + 6.992s^3}{0.833 + 1.012s + 2.439s^2 + 1.2s^3 + s^4}. \tag{16}$$

The complete tuning network for the antenna is shown in Fig. 15. Fig. 16 shows the frequency variation of TPG with and without the tuning network. Fig. 17 shows the variation of the effective radiative gain (=TPG$\times\eta_{\text{rad}}$) with frequency. It is seen that the present antenna radiates more power than the resistively loaded antenna or the unloaded conducting antenna over the desired band. However, since the radiation efficiency of the antenna is not uniform over the band (see Table. 5.), the effective radiative gain varies from about 5% to about 12.

To make the antenna truly broadband, we have attempted to design a matching network where the TPG is not constant but assumed to vary in such a fashion so as to make $TPG \times \eta_{\text{rad}}$ uniform over 30–90 MHz. We have used the software MATCHNET [9] for this purpose. Fig. 18 shows the topology of the designed matching networks. Two cases were considered: the first in which the capacitors and inductors were assumed lossless, and a second where lossy elements were allowed. All capacitors in the latter case had a Q factor of 500, while the inductors had a Q factor of 50. Fig. 19 shows the variation of effective radiative gain of the designed systems. By comparing with Fig. 17, we see that the overall gain is more uniform in the present case than in the latter case. The gain variation in the present case is from 6.7% to 10.5%. Furthermore, it is observed from Fig. 18 that the network does not have any transformers; a factor which is highly desirable from a practical standpoint. The networks of Fig. 18 are the ones that are recommended for the doubly loaded monopole.

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TABLE 1. LUMPED LOADS FOR 1 m MONOPOLE

n	z_n (m)	R_n (ohms)
1	1/9	20.21
2	2/9	22.90
3	3/9	26.46
4	4/9	31.28
5	5/9	38.28
6	6/9	49.36
7	7/9	69.58
8	8/9	118.94

$h = 1\,\text{m}, a = 0.5\,\text{cm}, N = 9$

TABLE 2. INPUT IMPEDANCE OF MONOPOLE

Freq. (MHz)	$Z_{\text{in}}^{\text{unloaded}}$ (Ohms)	$Z_{\text{in}}^{\text{loaded}}$ (Ohms)
30	$3.87 - j347.5$	$71.37 - j354.3$
40	$7.4 - j224.$	$80.58 - j235.$
50	$12.75 - j138.8$	$93.73 - j156.7$
60	$20.83 - j70.62$	$111.9 - j99.6$
70	$33.24 + j9.5$	$136.4 - j56.89$
80	$53.2 + j50.7$	$167.8 - j27.15$
90	$86.23 + j114.6$	$205.4 - j12.73$

$h = 1\,\text{m}, a = 0.5\,\text{cm}$

TABLE 3. RADIATION EFFICIENCY OF LOADED MONOPOLE

Freq. (MHz)	P_{in} (W)	P_{loss} (W)	η_{rad} (%)
30	2.44×10^{-4}	2.305×10^{-4}	5.54
40	5.56×10^{-4}	5.06×10^{-4}	8.98
50	1.16×10^{-3}	1.01×10^{-3}	12.92
60	2.12×10^{-3}	1.76×10^{-3}	17.14
70	2.99×10^{-3}	2.35×10^{-3}	21.35
80	3.12×10^{-3}	2.34×10^{-3}	21.12
90	2.75×10^{-3}	1.97×10^{-3}	28.36

$h = 1\,\text{m}, a = 0.5\,\text{cm}, V_{\text{in}} = 1\,\text{W}$

TABLE 4. STRAIGHT LINE AND RATIONAL RESISTANCE VALUES

k	ω_k	$R_q(\omega_k)$(Ohms)	$\tilde{R}_q(\omega_k)$ (Ohms)
0	0.0	0.	0.
1	1/9	19.65	21.03
2	2/9	88.77	87.50
3	3/9	191.06	194.06
4	4/9	312.5	301.97
5	5/9	364.13	375.53
6	6/9	428.63	431.36
7	7/9	525.23	512.74
8	8/9	646.80	653.12
9	1.0	726.97	725.59
10	1.5	0.0	19.53

$$\tilde{R}_q(\omega) = \frac{1667.4\omega^2}{1 - 1.93\omega^2 + 15.84\omega^4 - 19.53\omega^6 + 1.91\omega^8 + 5\omega^{10}}$$

TABLE 5. Input Impedance of Doubly Loaded Monopole

Freq. (MHz)	Z_1 (Ohms)	Z_2 (Ohms)	$Z_{\text{in}}^{\text{loaded}}$ (Ohms)	η_{rad} (%)
30	$60.44 + j79.5$	$19.1 + j50.0$	$21.6 - j309.7$	19.2
40	$107.4 + j78.7$	$40.9 + j66.8$	$50.1 - j180.2$	16.7
50	$149.3 + j48.5$	$77.5 + j75.0$	$95.2 - j107.1$	15.2
60	$165.0 + j0.74$	$123.8 + j57.0$	$135.3 - j86.0$	15.2
70	$154.1 - j41.0$	$149.7 + j6.7$	$141.5 - j85.1$	18.2
80	$131.5 - j66.4$	$136.5 - j42.9$	$131.7 - j76.8$	24.6
90	$108.4 - j78.3$	$106.9 - j67.9$	$130.5 - j37.4$	33.6

$h = 1\,\text{m}$, $a = 0.5\,\text{cm}$, $R_1 = 165\,\text{ohms}$, $L_1 = 0.5\mu H$, $C_1 = 14pF$, $R_2 = 150\,\text{ohms}$, $L_2 = 0.25\mu H$, $C_2 = 20pF$.

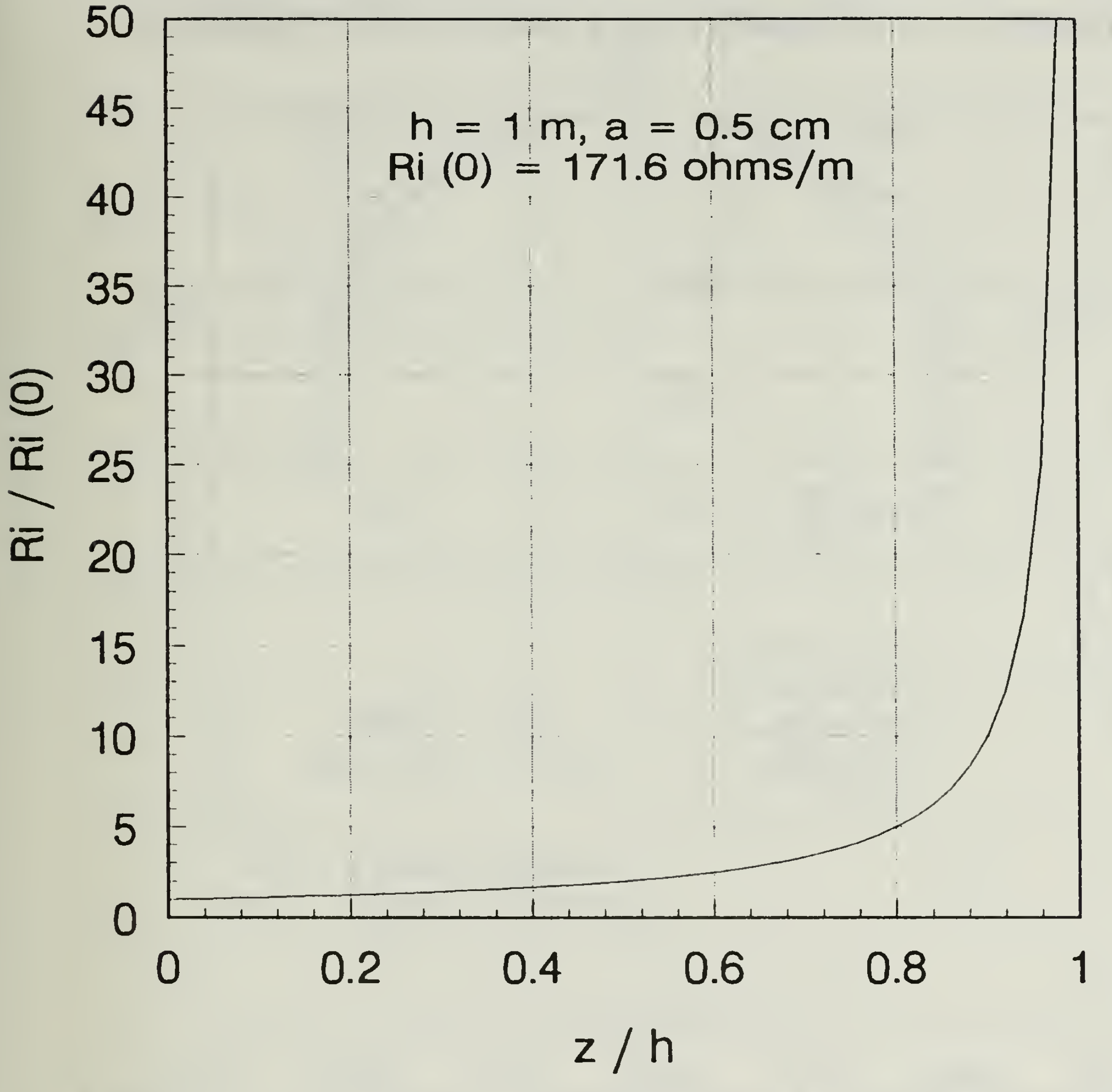


Fig. 1 Variation of Resistivity along Monopole

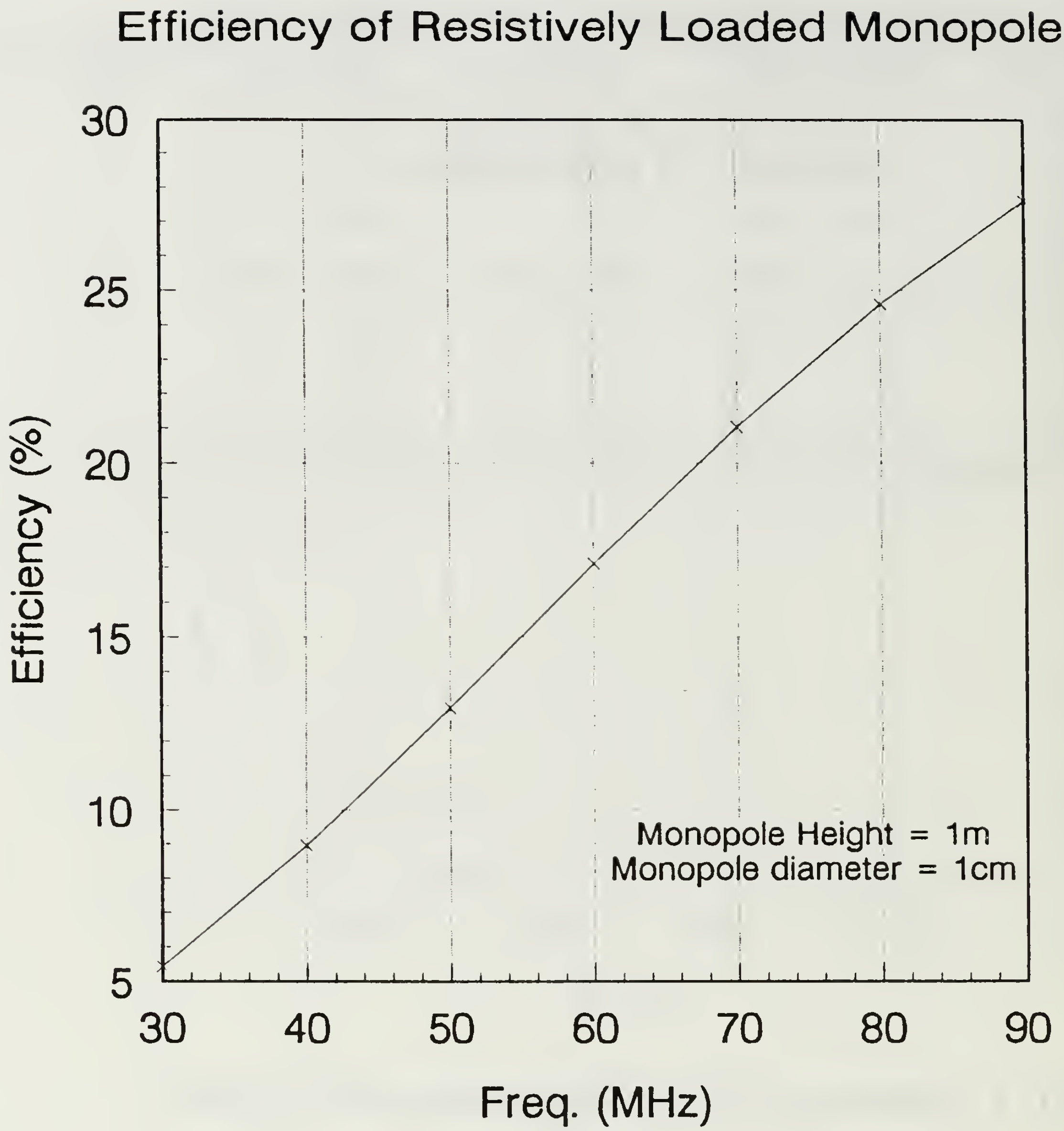
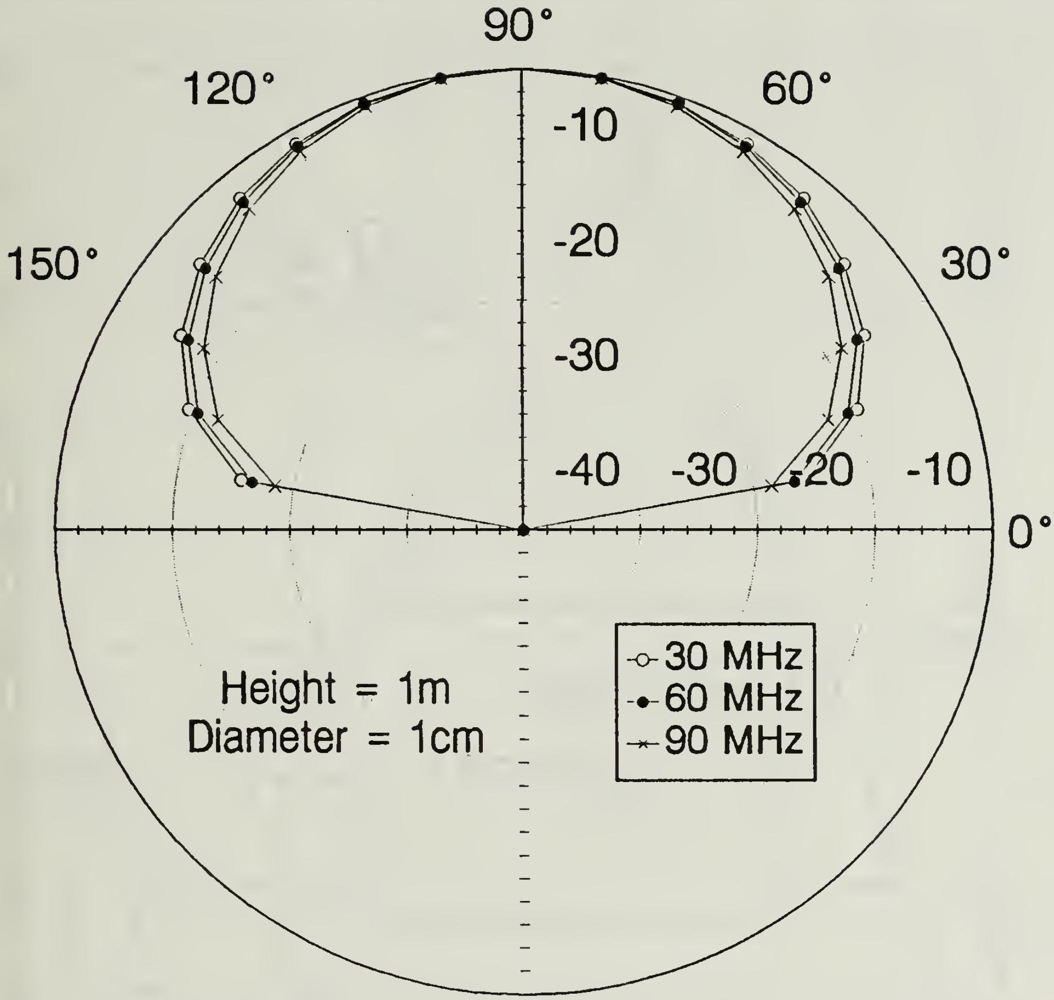

Efficiency of Resistively Loaded Monopole
Efficiency (%)
30
25
20
15
10
5
30
40
50
60
70
80
90
Freq. (MHz)
Monopole Height = 1m
Monopole diameter = 1cm

Fig. 2



Radiation Pattern of Resistively Loaded Monopole

Fig. 3

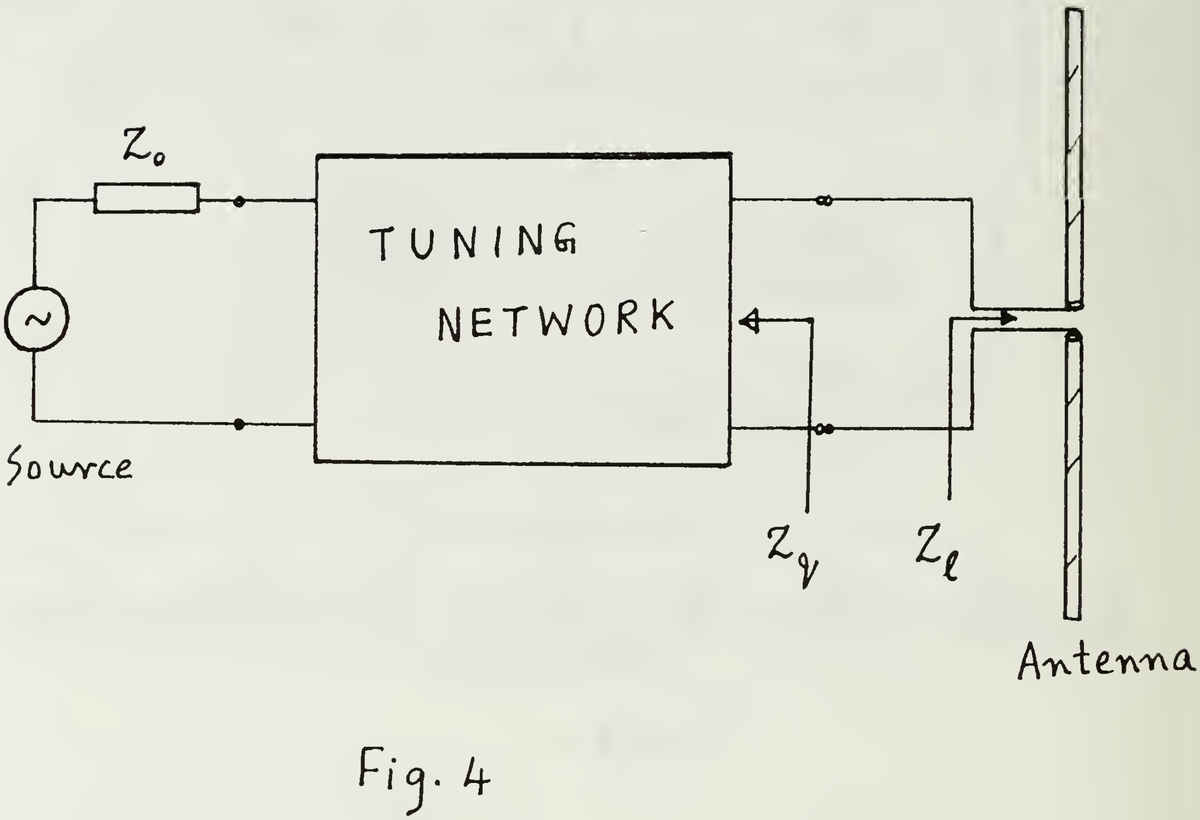
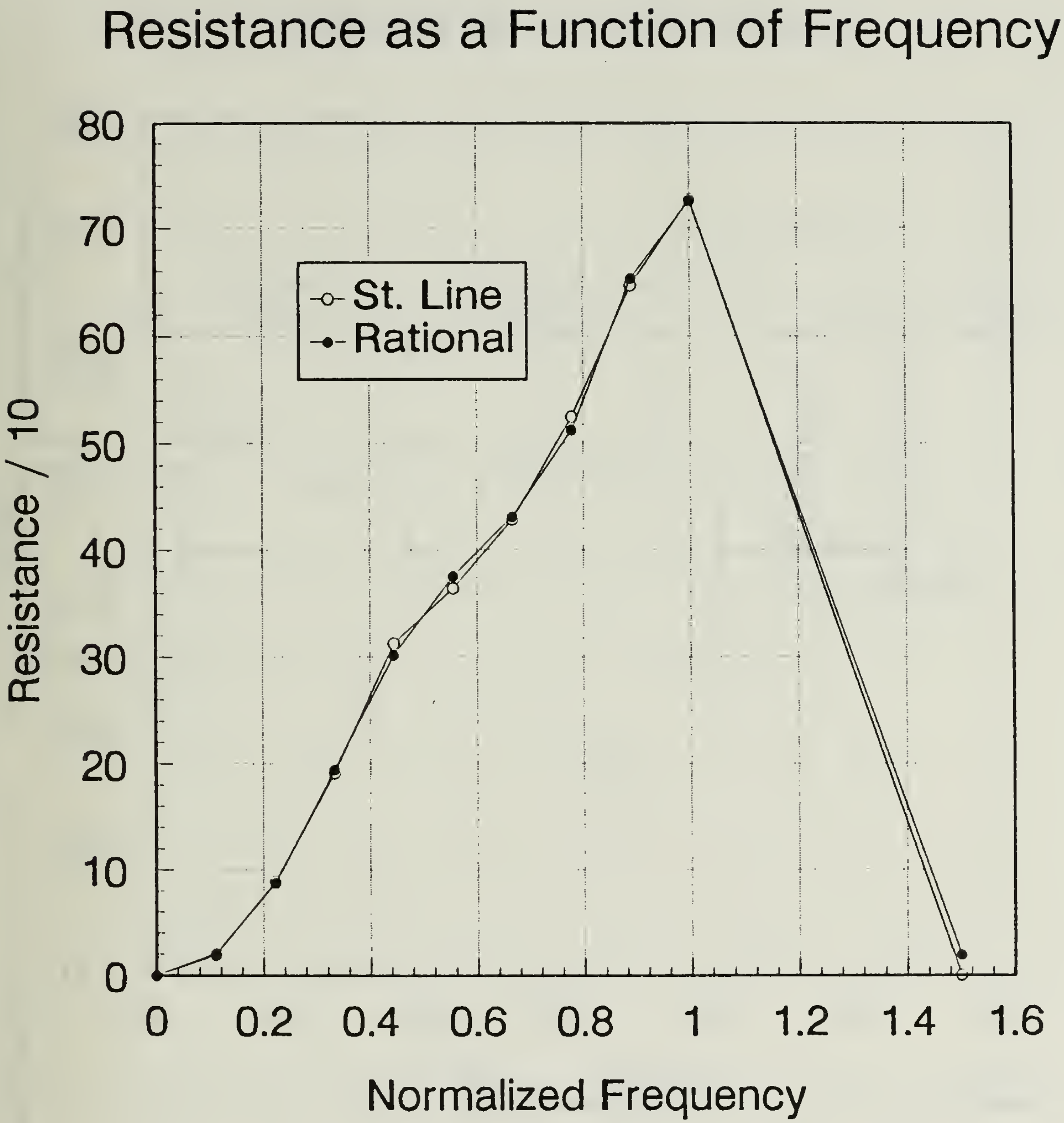


FIG.4. TUNING NETWORK FOR MATCHING ANTENNA TO SOURCE



Resistance as a Function of Frequency
80
70
60
50
40
30
20
10
0
Resistance / 10
St. Line
Rational
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
Normalized Frequency

Fig. 5

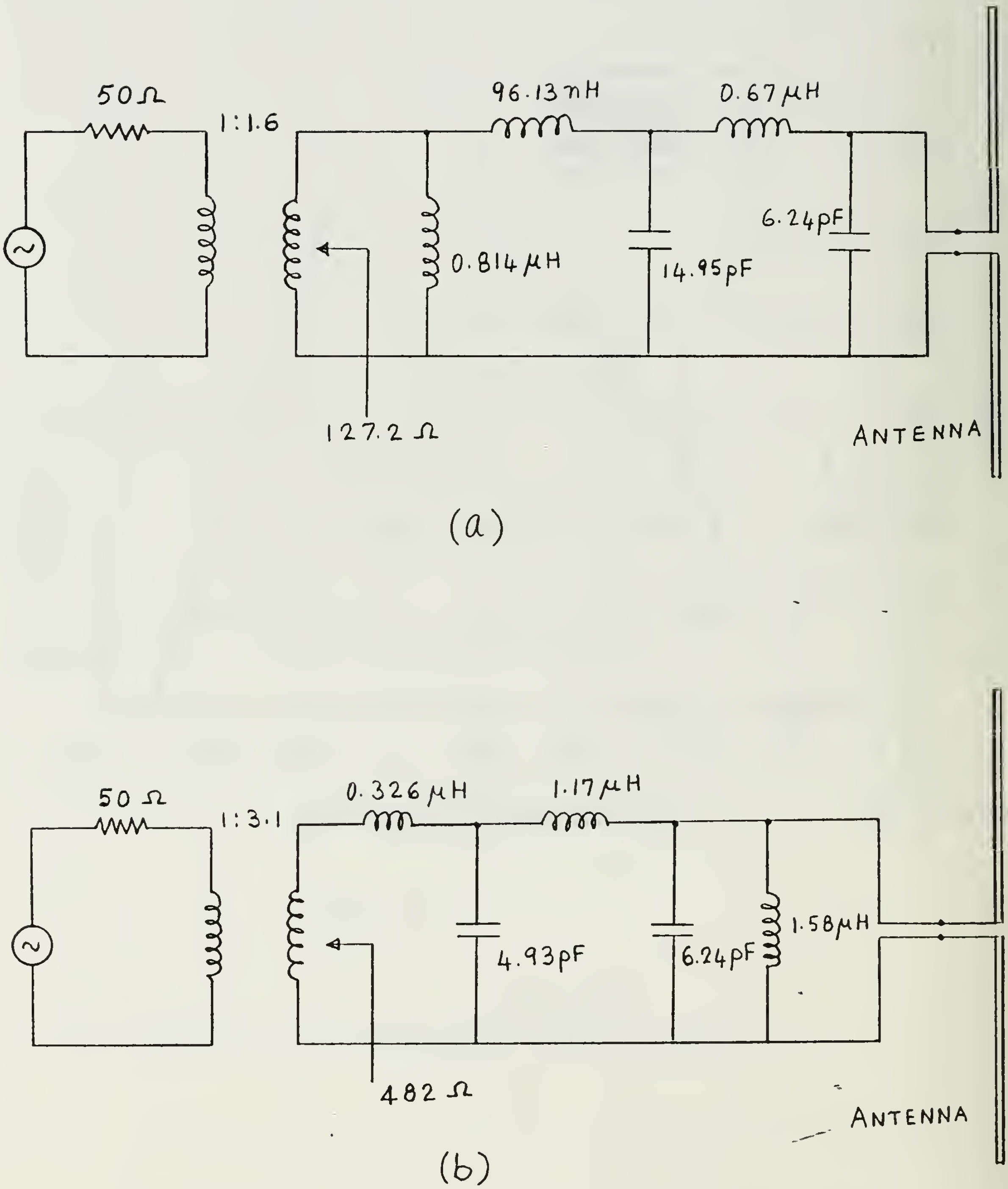


FIG. 6. CIRCUIT REALIZATIONS

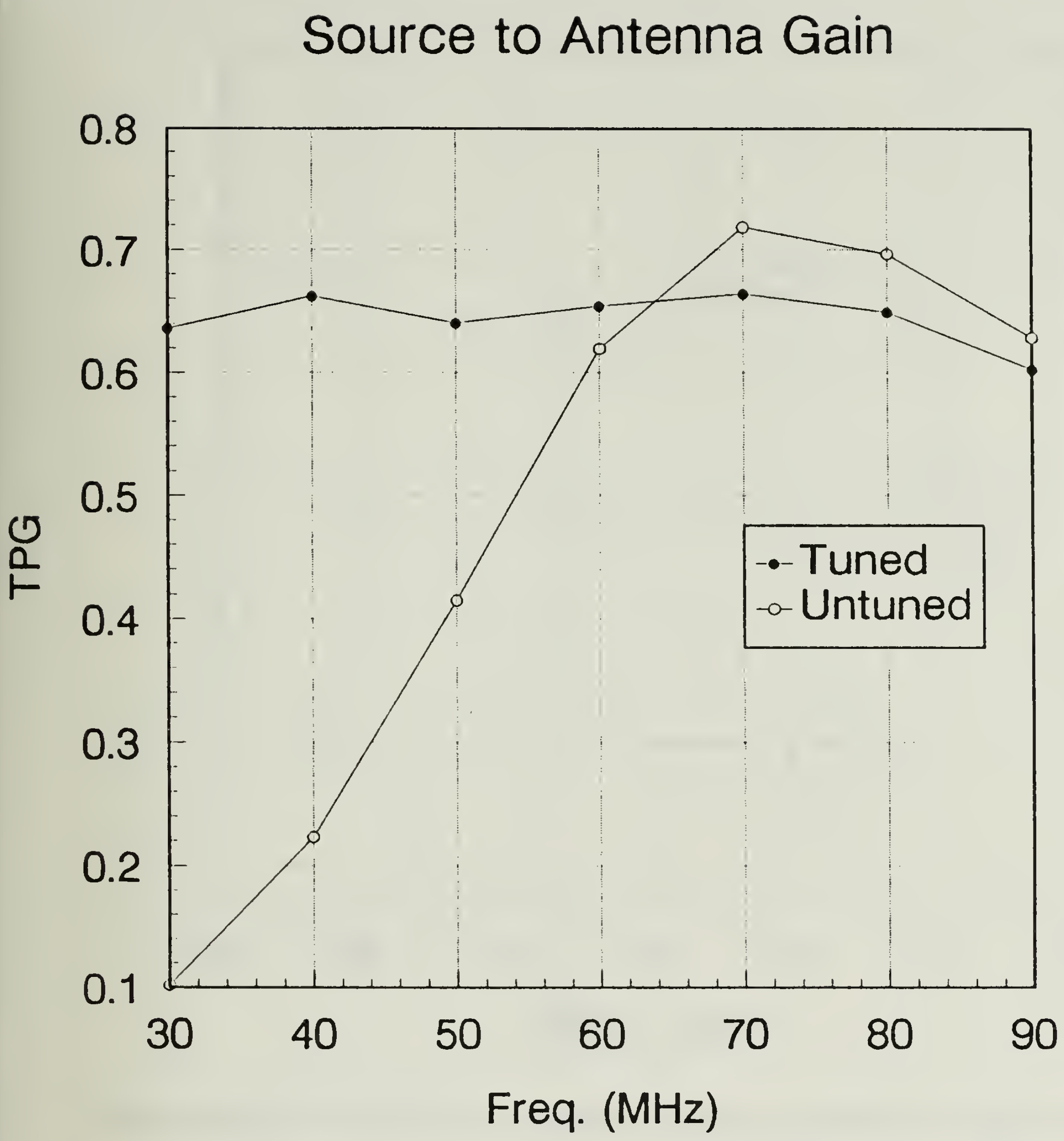


FIG. 7. TRANSDUCER POWER GAIN FOR RESISTIVELY LOADED MONOPOLE

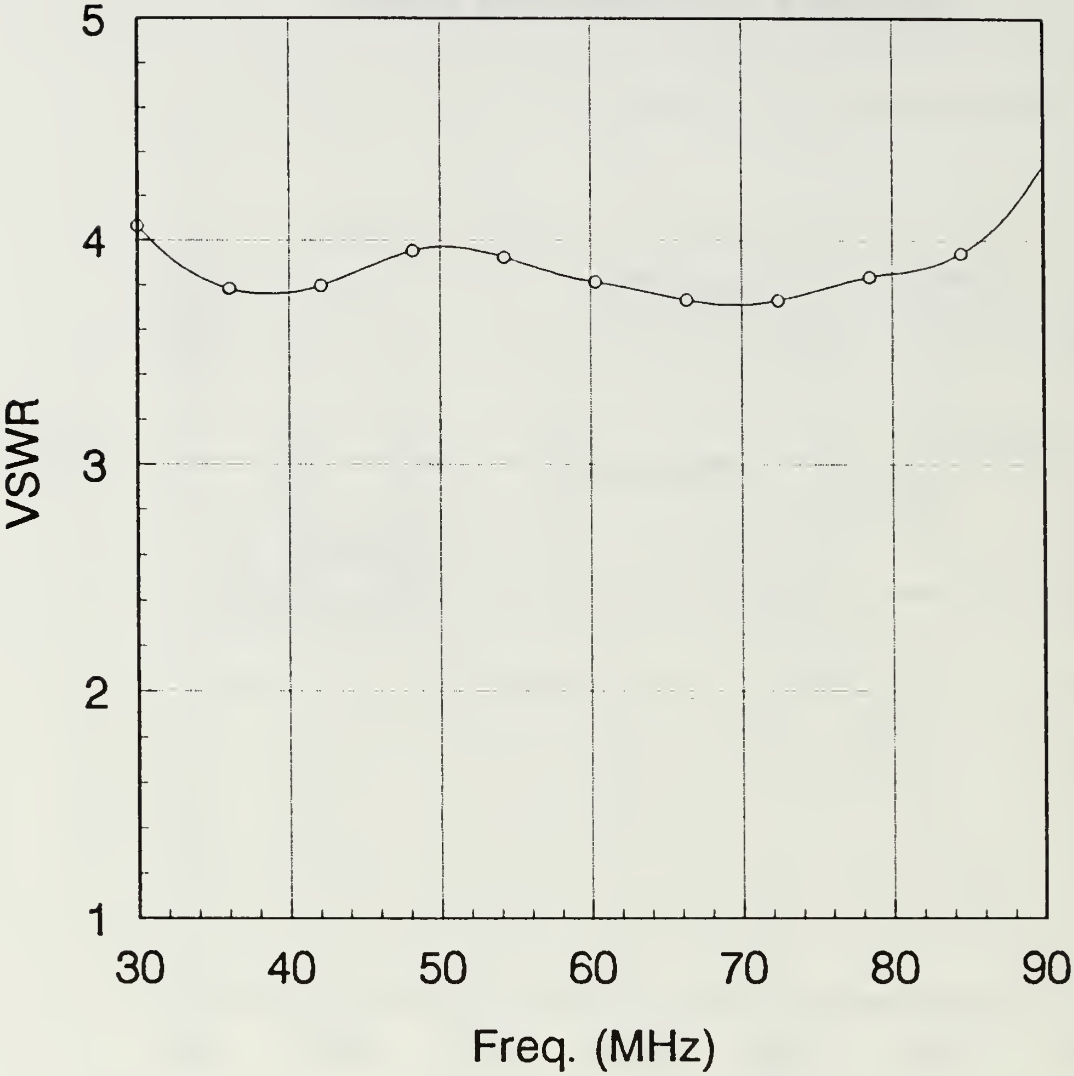


Fig. 8. Input VSWR as Seen From a 50 Ohm Generator

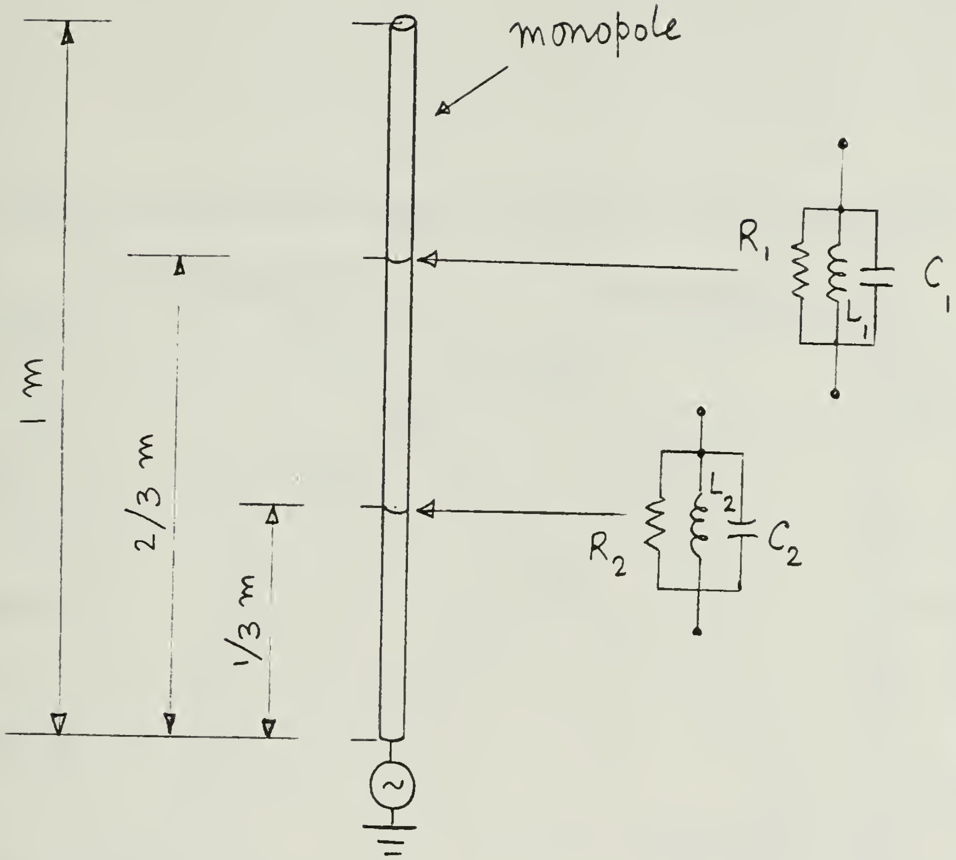


FIG. 9. Monopole Antenna with Double Loading

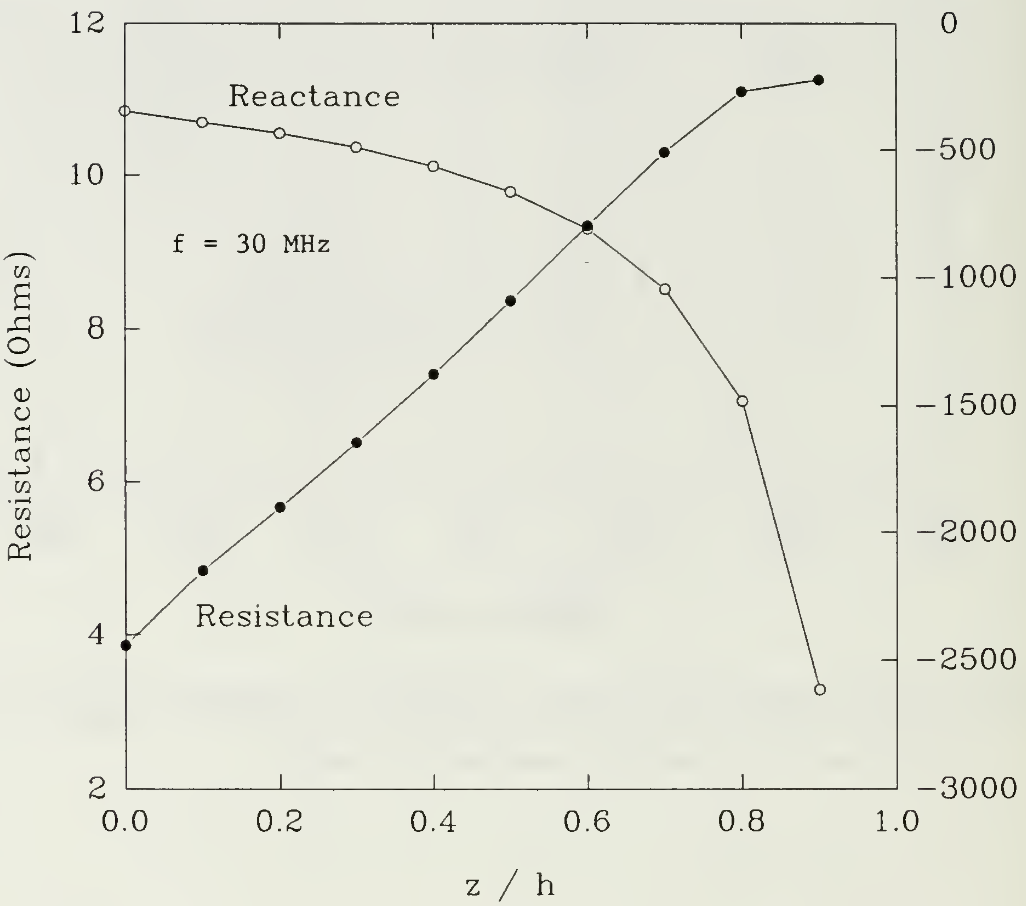


FIG. 10. INPUT IMPEDANCE OF 1m MONOPOLE VERSUS FEED LOCATION

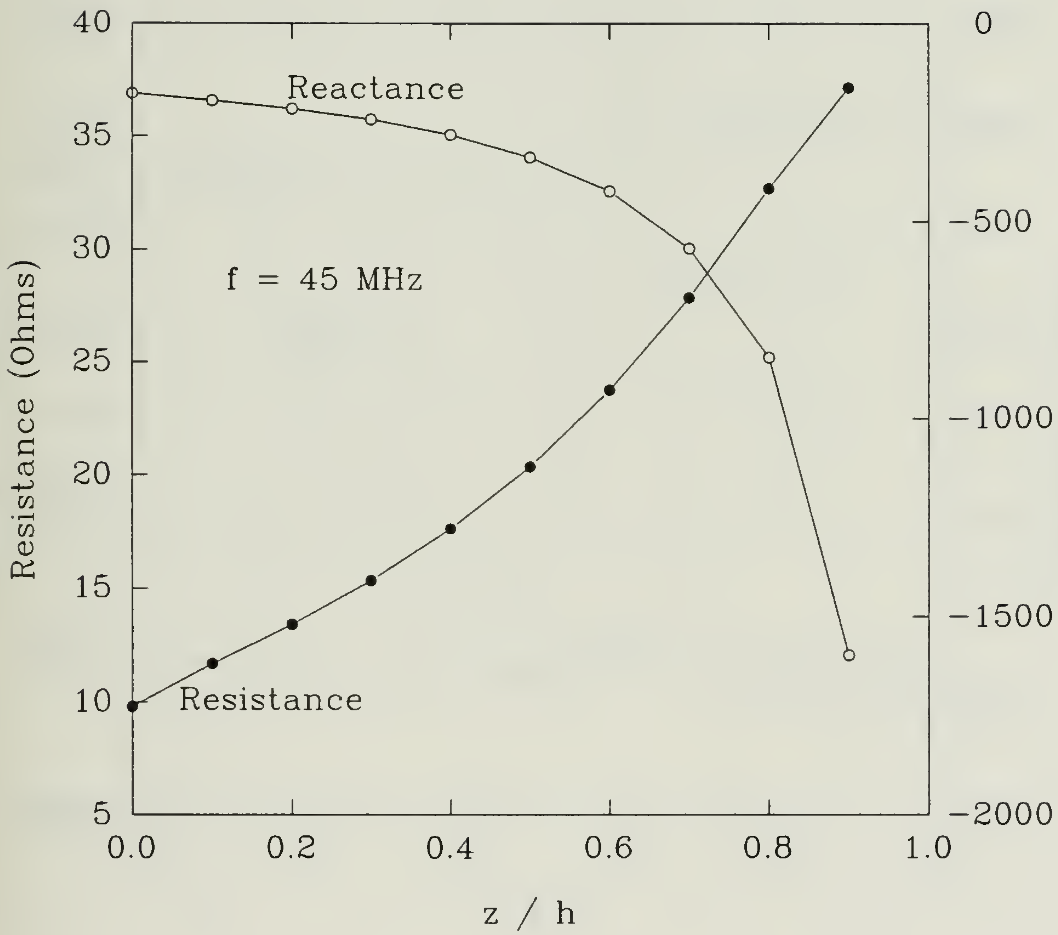


FIG. 11. INPUT IMPEDANCE OF 1m MONOPOLE VERSUS FEED LOCATION

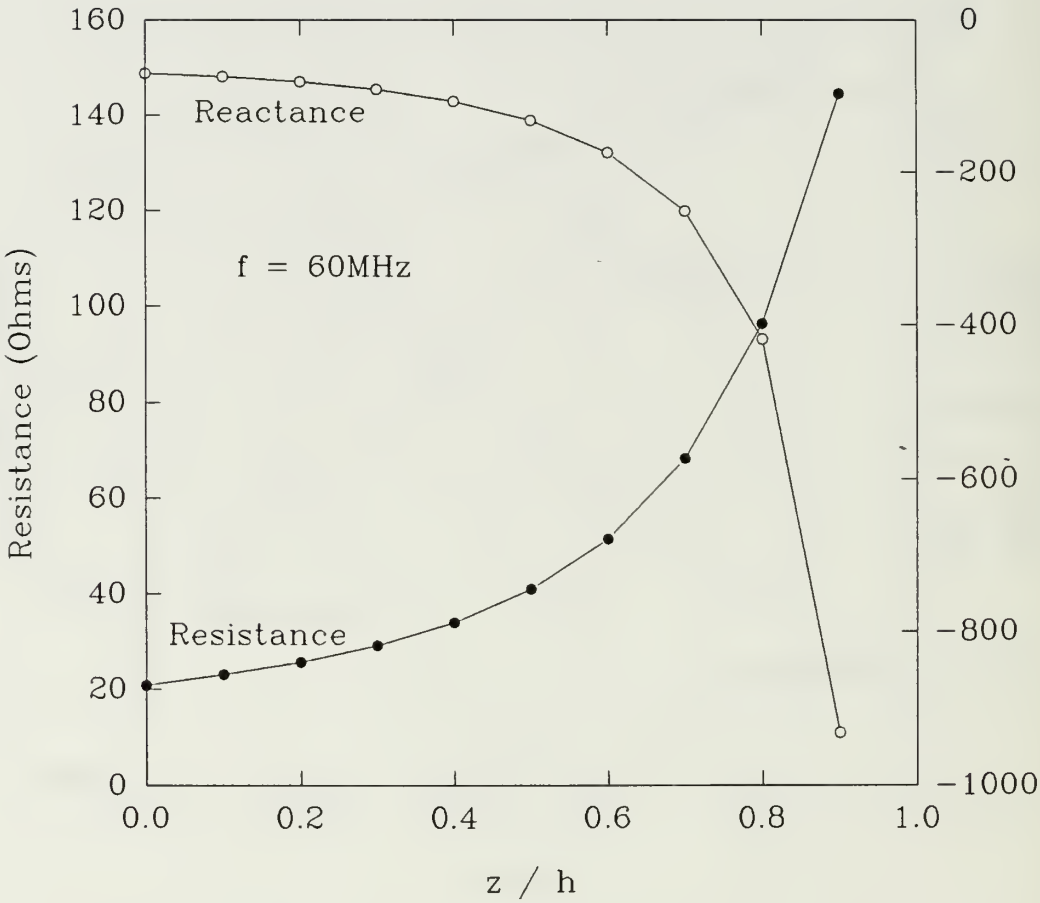


FIG. 12. INPUT IMPEDANCE OF 1m MONOPOLE VERSUS FEED LOCATION

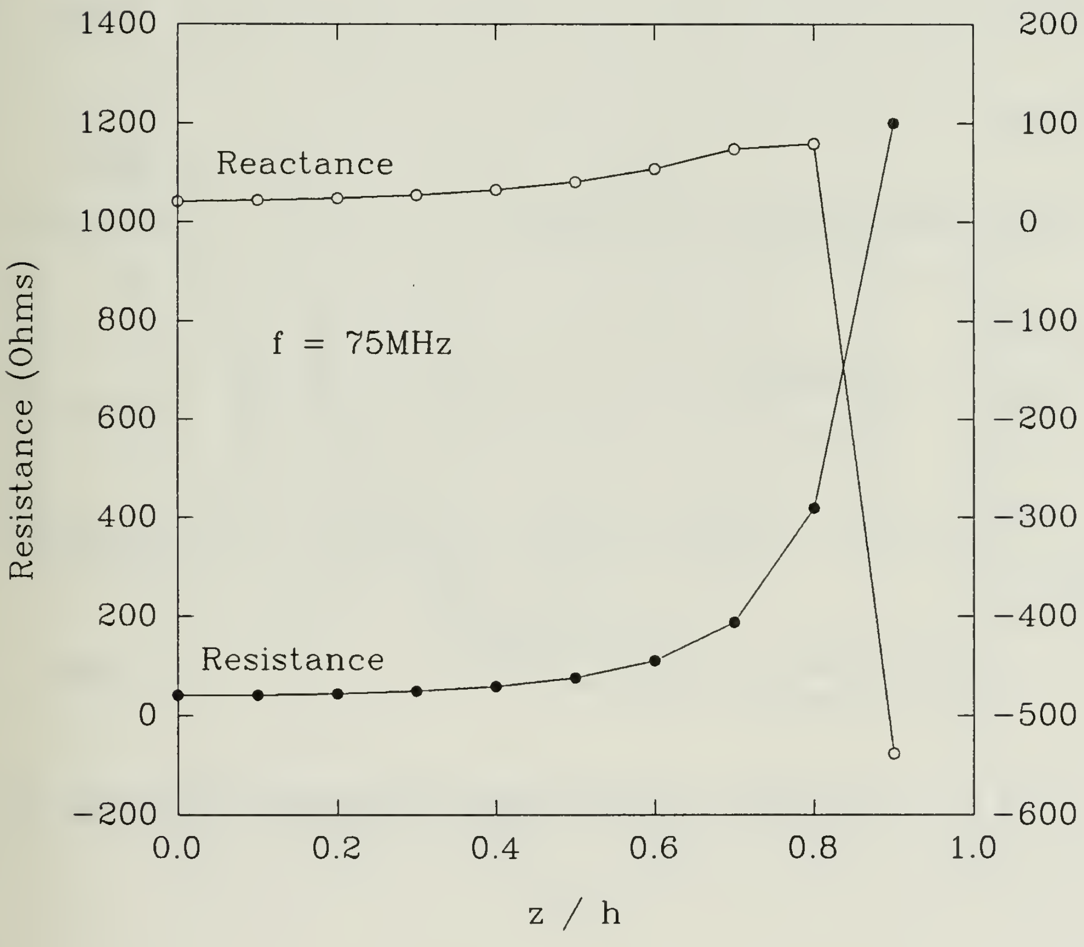


FIG. 13. INPUT IMPEDANCE OF 1m MONOPOLE VERSUS FEED LOCATION

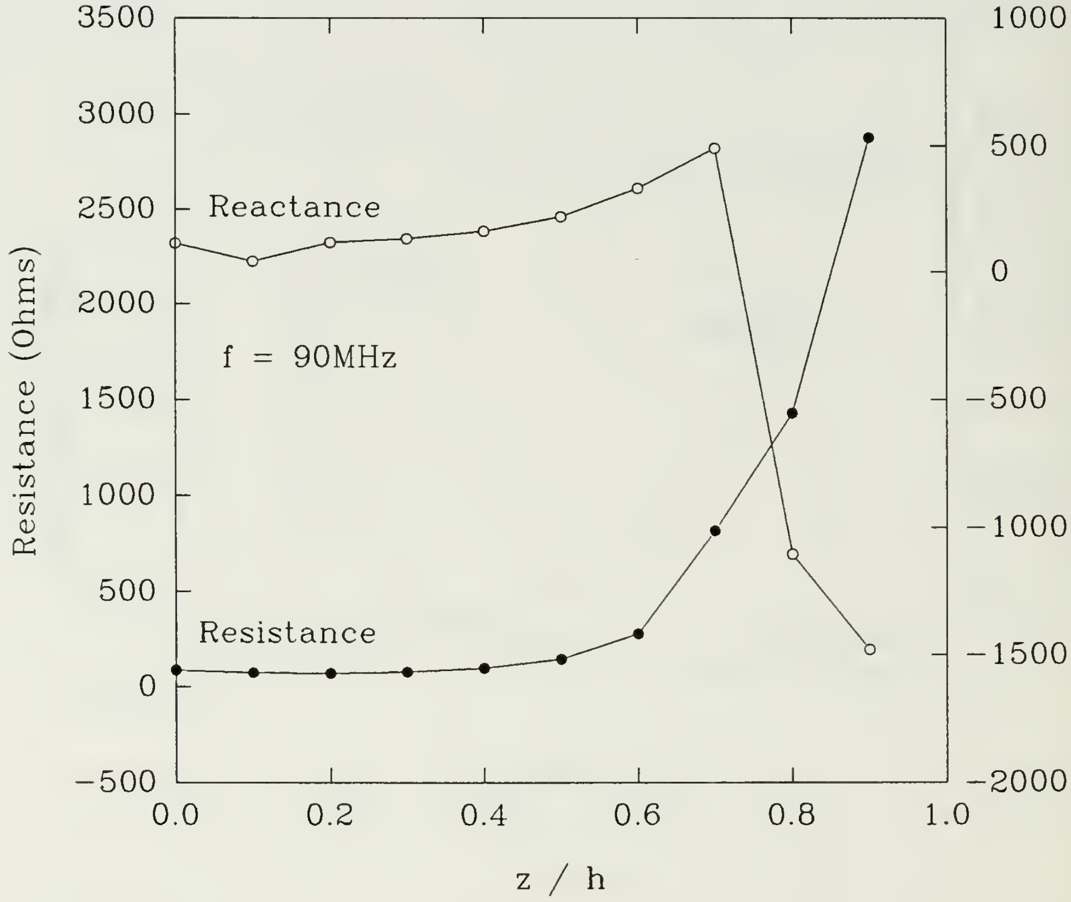


FIG. 14. INPUT IMPEDANCE OF 1m MONOPOLE VERSUS FEED LOCATION

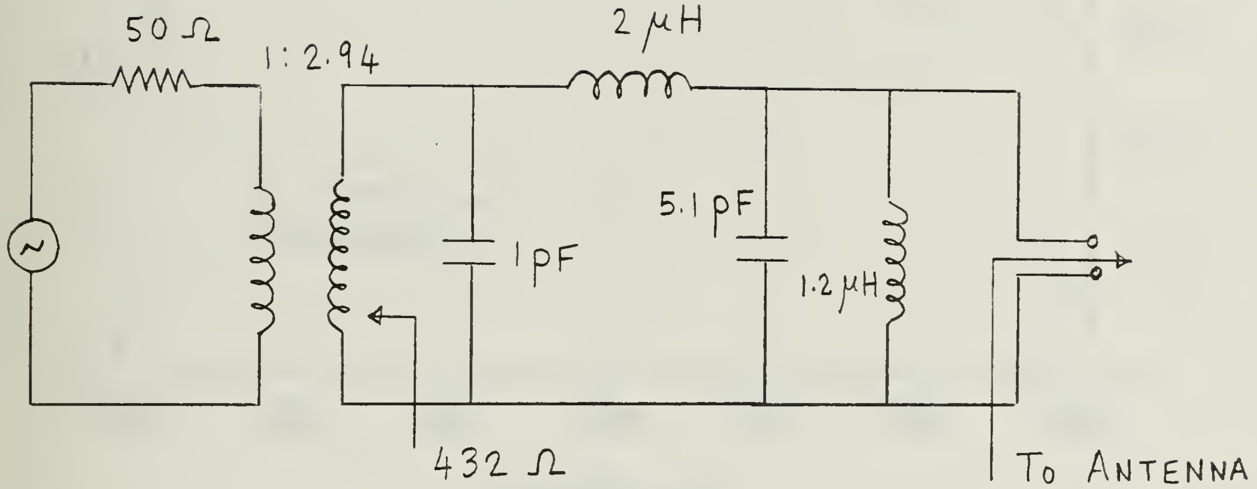


FIG. 15. TUNING NETWORK FOR DOUBLY LOADED 1m MONOPOLE

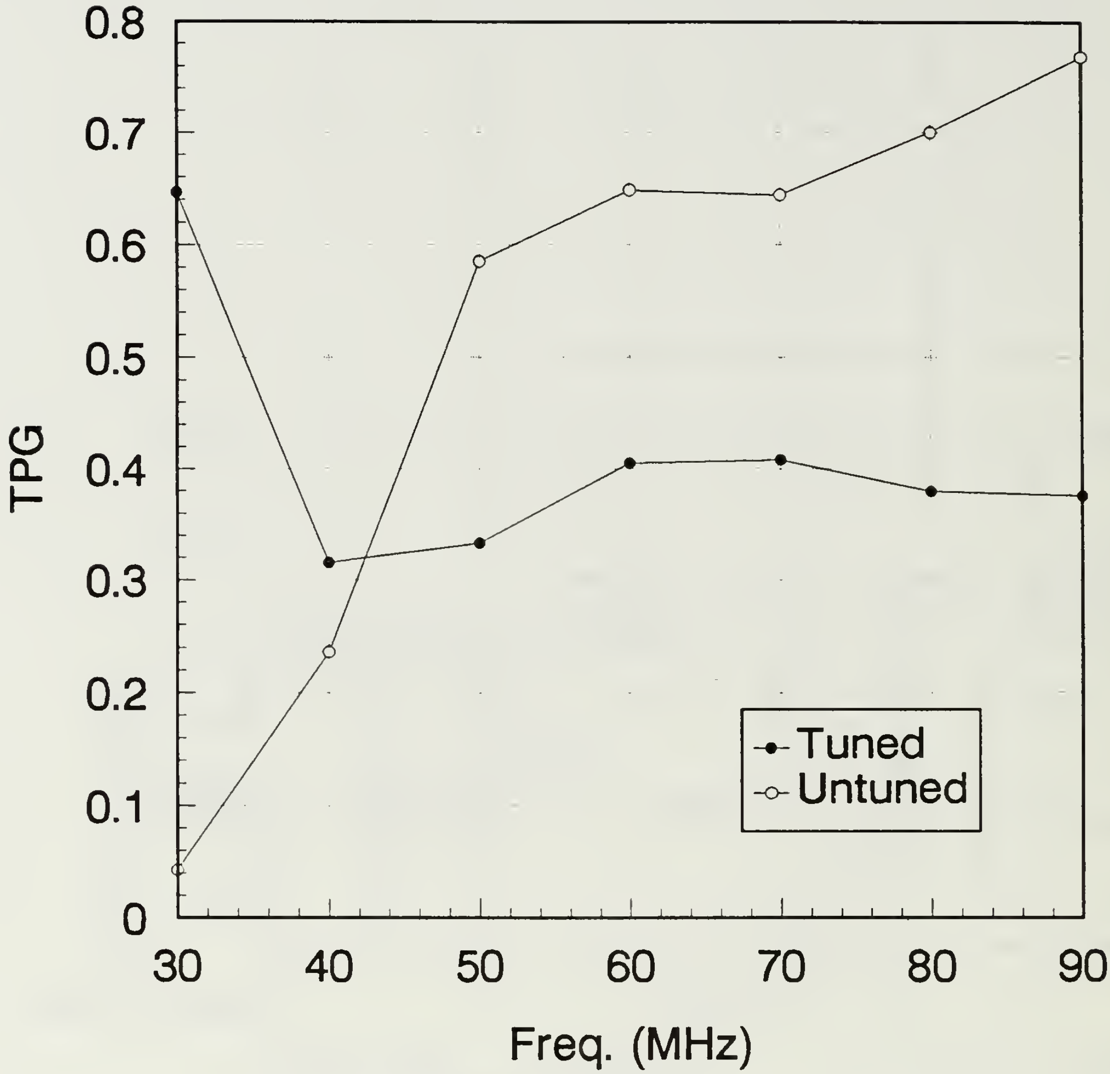


FIG. 16. TRANSDUCER POWER GAIN FOR DOUBLY LOADED MONOPOLE

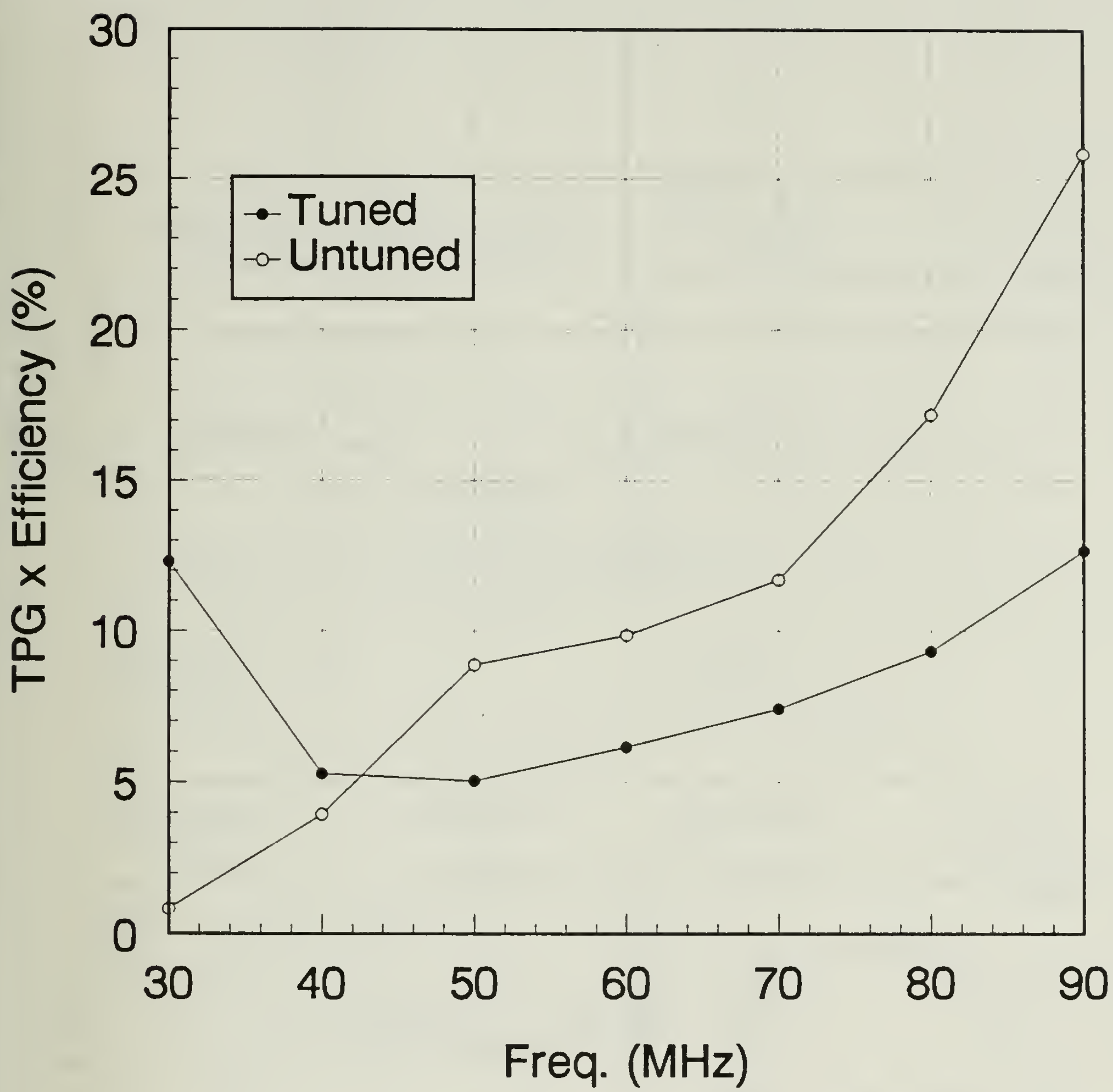


FIG. 17. EFFECTIVE RADIATIVE POWER GAIN FOR DOUBLY LOADED MONOPOLE

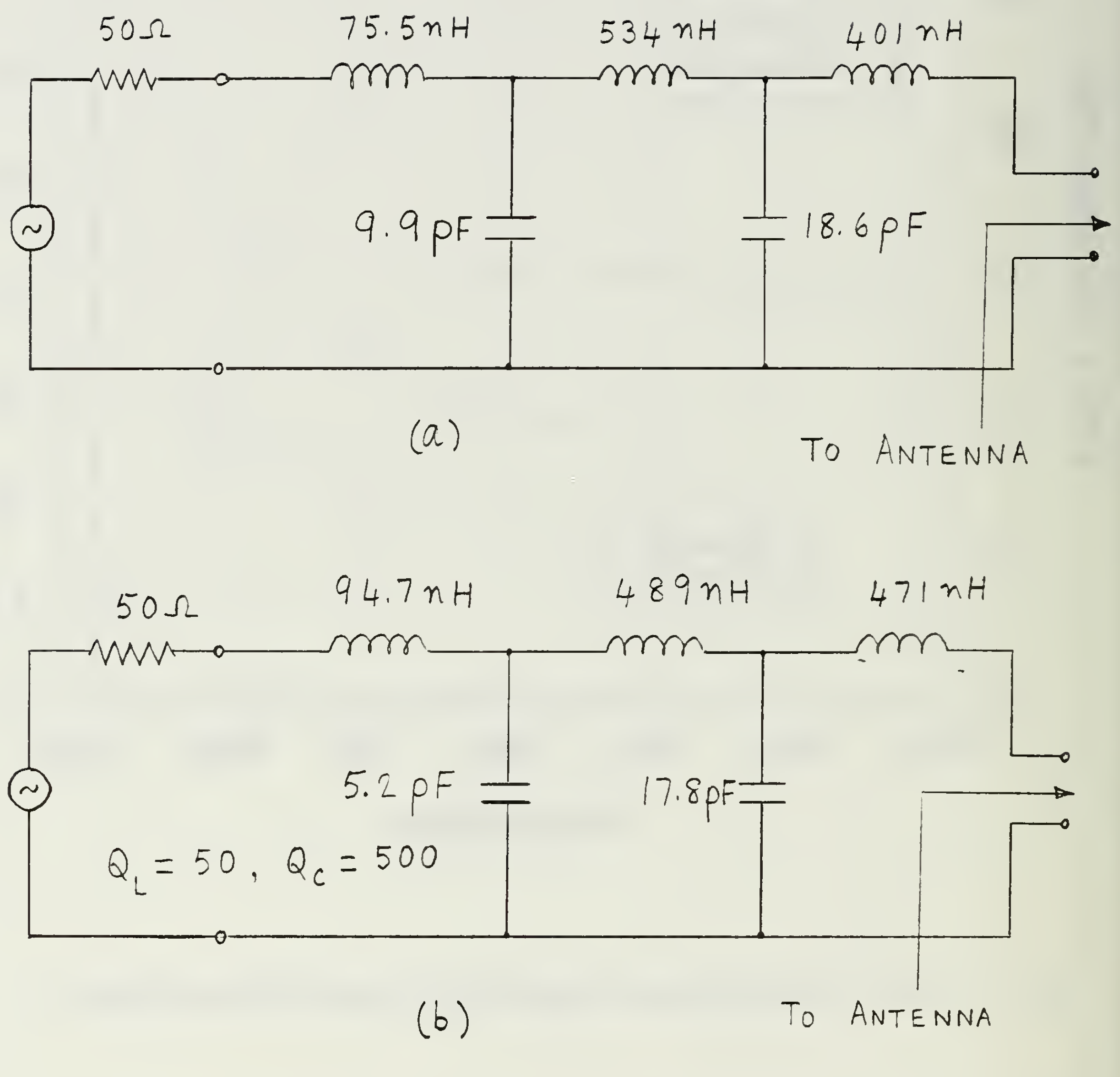


Fig. 18. Tuning Networks for Doubly Loaded Monopole. (a) Lossless elements (b) Lossy elements: $Q_L = 50$, $Q_C = 500$.

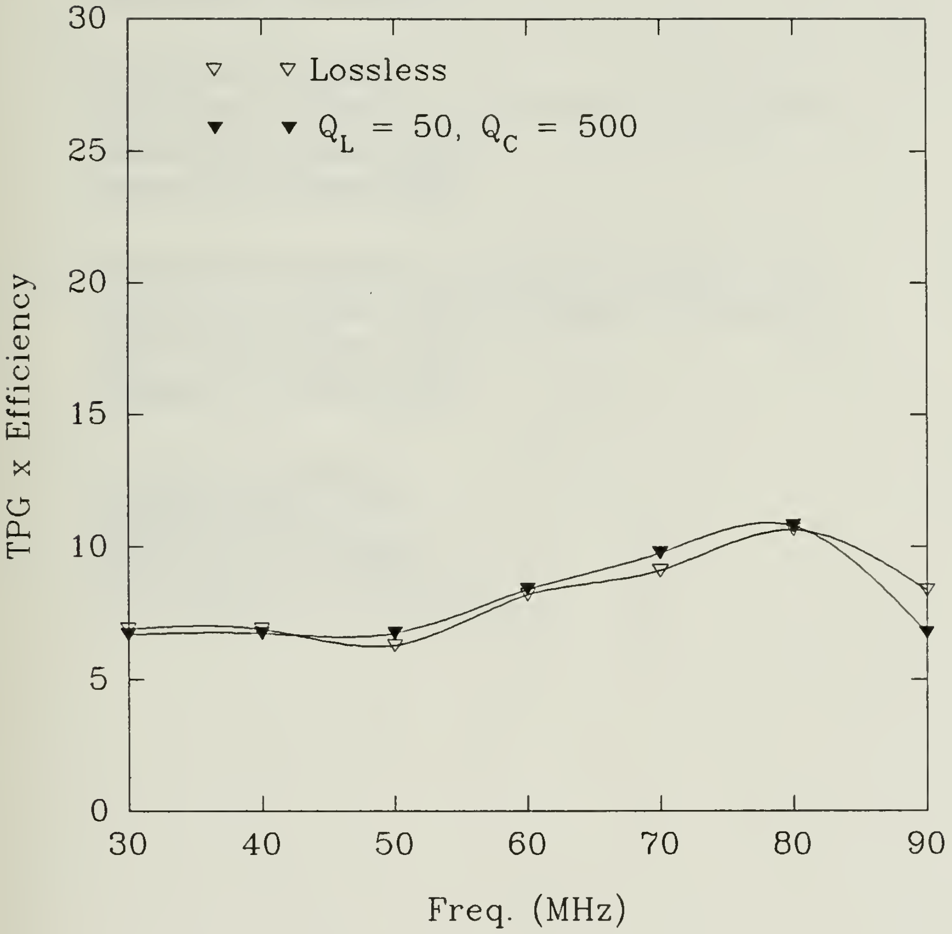


FIG. 19. Effective Radiative Gains with Tuning Networks of Fig. 18.

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